

PATENT APPLICATION

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Quadrature Product Subcarrier Modulation System

SPECIFICATION

Statement of Government Interest

The invention was made with Government support under contract No. F04701-93-C-0094 by the Department of the Air Force. The Government has certain rights in the invention.

Reference to Related Application

The present application is related to applicant's copending application entitled Coherent Adaptive Subcarrier Modulation Method, S/N: xx/xxx,xxx, filed yy/yy/yy, by one of the named inventors.

Field of the Invention

The invention relates to the field of modulation of signals of communication systems. More particularly, the present invention relates to subcarrier modulation methods for efficient frequency reuse within exiting bandwidths.

Background of the Invention

Due to the limited availability of spectrum allocations for communications systems, it has become desirable to reuse existing spectrum by employing bandwidth efficient modulation methods to existing satellite communications systems. In particular, it has become desirable to add new signals to existing quadrature multiplexed spread-spectrum communication signals, such as the signals employed in the Global Positioning System (GPS), within the existing spectral allocation. As the number of users of a given communication system increases, it is often desirable to augment the system with additional communication signals. One method that has been used to achieve this goal in timing, telemetry and command links is conventional subcarrier phase modulation. Furthermore, it is often necessary to modulate new signals onto existing quadrature multiplexed communication systems within the existing spectral allocation. It is desirable that any such approach satisfy various constraints, including causing minimal distortion of existing signals, transmitting new signals through the same high power amplifier used by existing signals, accommodating new data messages and new pseudo random noise (PRN) code families for spread-spectrum systems, providing flexibility to control the spectral separation of signals within the allocated band, and providing flexibility to control the distribution of energy in and outside of the allocated band. Unfortunately, for an existing quadrature-multiplexed communication systems having data modulations on both the I and Q channels, that is, the in-phase channel with zero degree phase

1 baseline Block IIF satellite will also transmit these signals on
2 both the GPS L1 and L2 frequencies, along with the new military
3 signals.

4
5 The new M-code signal consists of the product of a military
6 data modulation multiplied by a spreading code modulation.
7 Previously proposed approaches for augmenting the existing GPS
8 waveform, with a new split-spectrum M-code signal, such as the
9 tri-code hexaphase modulation method, involve the linear addition
10 of a single new M-code signal on to the existing GPS waveform.
11 These approaches multiplex the additional signal with an existing
12 I/Q quadrature modulation by adding a third signal to one of the
13 I and Q phases. Unfortunately, this gives rise to the
14 undesirable non-constant envelope, that is, a variable amplitude.
15 For this case, the envelope of the composite signal is no longer
16 constant due to the presence of the time-varying amplitude
17 component in addition to the constant component of the envelope.
18 This result is undesirable because the amplitude variations will
19 give rise to AM-to-AM and AM-to-PM distortions when the signal is
20 passed through a nonlinear amplifier unless the operating point
21 of the amplifier is backed off from its saturation point to the
22 linear region of amplifier operation. Such a back off can result
23 in appreciable power losses.

24
25 During the course of the GPS military signal development
26 study, a number of M-code signal modulation methods have been
27 proposed. Two of the leading approaches were hard-limiting the
28 sum of the C/A, P and M-code signals, and majority voting. The

1 hard-limiting approach involves forcing the non-constant envelope
2 waveform resulting from the sum of C/A and M on one carrier phase
3 and P on the other, to be constant by limiting the amplitude
4 variation to a minimum value. This technique results in
5 significant signal power loss and distortion for the case of
6 equal C/A and M-code power levels as is the case for hard-limited
7 tri-code hexaphase modulation method. Furthermore, the exact
8 efficiency is critically related to the power balance between
9 C/A, P(Y) and M-code signals and the desired balance between
10 signals is not easy to achieve. An alternative approach to
11 combining M-code with C/A and P(Y) code signals is through
12 majority vote combining. In the majority vote approach, signals
13 are time multiplexed, that is, time-shared, on either I or Q
14 phases to allow multiple signals to be transmitted in a single
15 constant envelope. The disadvantage of this approach results
16 from the relatively large majority combining loss per code in
17 combination, assuming equal power levels for all codes in the
18 majority combination. Furthermore, it is difficult to control
19 the relative power levels of the combined signals without
20 incurring additional combining losses.

21
22 Conventional subcarrier modulation has been recognized as a
23 means to modulate additional signals onto a modulated carrier
24 signal while maintaining a constant envelope. The
25 Space-Ground-Link Subsystem for example, employs three
26 subcarriers that are phase-modulated onto a carrier signal to
27 enable the modulation of four signals, including one carrier
28 modulation and three subcarrier modulations. The classical

1 The leading modulation methods, that have been proposed thus
2 far to achieve this end, have a maximum efficiency limited to
3 roughly seventy-one percent in the allocated GPS spectral band.
4 Efficiency is defined as the sum of the effective transmitter
5 power after reception plus any band limiting losses and other
6 losses associated with this sum divided by the total transmitted
7 power. At the same time, the leading approaches with the highest
8 efficiency restrict options available to optimize the waveform by
9 transmitting a single M-code signal used for both acquisition and
10 tracking.

11
12 These single M-code approaches include tri-code hexaphase,
13 hard-limited tri-code hexaphase, majority vote code multiplexing,
14 and offset carrier and binary offset carrier modulations. The
15 tri-code hexaphase, offset carrier and binary offset carrier
16 modulation approaches have a non-constant envelope. As a result,
17 these approaches cause distortion to existing signals when the
18 total waveform is passed through a non-linear amplifier. The
19 majority vote modulation approach employs a constant envelope but
20 does not provide significant spectral separation and has limited
21 inherent flexibility in adjusting the amplitude of generated
22 harmonics. The majority vote modulation approach also suffers
23 from majority combining losses that results in relatively poor
24 overall power efficiency. The same is true for the hard-limited
25 tri-code hexaphase modulation method.

26
27 The transmission of both data and spreading code modulations
28 in a quadrature-multiplexed communication channel is well known.

1 Conventional quadrature phase shift keying (QPSK) direct-sequence
2 spread spectrum modulation schemes generally employ spreading
3 sequences on the in-phase and quadrature channels. The
4 modulation of unspread data is a special case. For GPS L1 or L2
5 carriers modulated by codes and data, a carrier signal is
6 modulated by codes and data in a conventional quadrature
7 modulated communication signal as $S_0(t)$ in terms of the in-phase
8 $I_0(t)$ and quadrature $Q_0(t)$ signals. The signal $S_0(t)$ can express
9 GPS L1 and L2 carriers modulated by C/A and P codes and data in a
10 conventional quadrature modulation communication method in terms
11 of an $S_0(t)Q_0/I_0$ equation with respect to an I_0 equation and a Q_0
12 equation.

$$S_0(t) = I_0(t) \cos(\omega_c t) - Q_0(t) \sin(\omega_c t)$$

$$I_0(t) = \sqrt{\frac{2P_I}{P_T}} D_I(t) C_I(t)$$

$$Q_0(t) = \sqrt{\frac{2P_Q}{P_T}} D_Q(t) C_Q(t)$$

The $C_I(t)$ and $D_I(t)$ terms are the in-phase pseudo random noise spreading code and data modulation, and $C_Q(t)$ and $D_Q(t)$ are the quadrature pseudo random noise spreading code and data modulations. P_I and P_Q are the average power of the I and Q signals in relation to total power P_T .

The signal $S_0(t)$ may also be expressed in an $S_0(t)A_0/\phi$ equation. An $A_0(t)$ magnitude equation describes the envelope form where $A_0(t)$ is the magnitude envelope of $S_0(t)$. A $\phi(t)$ equation describes the phase of $S_0(t)$.

$$S_0(t) = A_0(t) \cos(w_c t + \phi(t))$$

$$A_0(t) = \sqrt{I_0^2(t) + Q_0^2(t)} = \sqrt{\frac{2(P_I + P_Q)}{P_T}} = \text{CONSTANT}$$

$$\phi(t) = \tan^{-1} \frac{Q_0(t)}{I_0(t)} = \tan^{-1} \sqrt{\frac{P_Q}{P_I} \frac{D_Q(t) C_Q(t)}{D_I(t) C_I(t)}}$$

The total average power may be described by a P_{TF} equation for the signal $S_0(t)$ that can be expressed by the magnitude of the envelope squared divided by two, in reference to identity equations based on the square of the C_I and C_Q code signals and the D_I and D_Q data signals.

$$P_{TF} = \frac{A_0^2}{2} = \frac{I_0^2 + Q_0^2}{2} = \frac{P_I + P_Q}{P_T}$$

$$C_I(t)^2 = C_Q(t)^2 = D_I(t)^2 = D_Q(t)^2 = 1$$

Conventional approaches to multiplexing additional signals with an existing I/Q quadrature modulation involve adding a third signal to one of the two I or Q phases. Unfortunately, this gives rise to the undesirable non-constant envelope characteristic. A signal can be linearly added to the I phase of the baseband waveform modulated onto the same carrier as the existing I/Q signals. For this case, the magnitude envelope of the signal $A_0(t)$ is no longer constant due to presence of the time-varying component of $A(t)$ in addition to the constant component of $A_0(t)$. This result is undesirable because the amplitude variations will give rise to amplitude modulation to amplitude modulation (AM-to-AM), and amplitude modulation to phase modulation (AM-to-PH) distortions when the signal is passed through a nonlinear amplifier. These and other disadvantages are solved or reduced using the invention.

Summary of the Invention

An object of the invention is to provide a constant envelope composite signal with added subcarrier signals that do not distort existing signals when the composite signal is passed through a high-power amplifier operating near saturation.

Another object of the invention to provide for the modulation of new signals onto a quadrature multiplexed communication channel while controlling the power balance between signals while maintaining a constant envelope.

Yet another object of the invention is control the spectral separation of quadrature multiplexed signals within the allocated band through the use of subcarrier frequencies, subcarrier codes rate, and a subcarrier modulation index.

Yet another object of the invention is to enable the modulation of an orthogonal pair of subcarrier signals on to the I and Q phases of a quadrature modulated carrier.

Another object of the invention is to transmit the new M-code signals using either separate or combined-apertures with reduced distortion and losses.

A further object of the invention is to provide control of the distribution of energy among signals inside and outside an allocated band.

1 Yet a further object of the invention is to utilize cross-
2 product intermodulation signals as useful communication signals.

3
4 In a general aspect of the invention, quadrature product
5 subcarrier modulation (QPSM) enables the transmission of a
6 quadrature multiplexed carrier modulation with one or more
7 subcarrier signals in the same constant-envelope waveform. The
8 generalized QPSM enables the application of subcarrier modulation
9 to quadrature multiplexed communication systems such as
10 quadrature phase shift keying (QPSK) or minimum shift keying
11 (MSK). The QPSM can be applied to both direct and
12 spread-spectrum quadrature multiplexed communication systems. In
13 particular, QPSM is advantageous for any spread spectrum system
14 desiring additional spread signals with spectral isolation
15 between new and existing pseudo random noise (PRN) code signals
16 using the same transmitter power amplifier. QPSM can augment
17 existing two-code spread-spectrum systems, without the need to
18 employ time multiplexing or majority voting that result in
19 significant power losses, while maintaining a constant envelope
20 signal with spectral separation between existing signals and new
21 signals with high efficiency. The QPSM enables a quadrature-
22 multiplexed subcarrier spread-spectrum waveform modulation using,
23 in general, multiple rate product codes that cause minimal
24 interference to existing codes and the new codes. The modulation
25 index can be used to control the distribution of energy between
26 carrier and subcarrier signals.

1 In a particular aspect of the invention, coherent adaptive
2 subcarrier modulation (CASM) is provided as a flexible, efficient
3 GPS modulation approach that can be tailored to different
4 civilian and military GPS modes of operation to provide high
5 efficiency without altering the basic modulation architecture.
6 This efficiency can be tailored using a suitable modulation
7 index, spreading code and subcarrier rates. CASM offers a
8 constant envelope subcarrier modulation method particularly
9 suitable the GPS. CASM provides a high efficiency of greater
10 than 90% with inherent flexibility to fine tune the modulation
11 architecture and maintain an ability to provide backward
12 compatibility with current GPS signals within the allocated 24.0
13 MHz GPS band. CASM can generate new GPS acquisition and tracking
14 codes from products of the current GPS ranging signals and a
15 newly defined subcarrier code and code partition function. CASM
16 can generate a military GPS acquisition and a military GPS
17 tracking signal from a single new code.

18
19 CASM is applicable to the transmission of quadrature
20 duplexed military acquisition (MA) and military tracking (MT)
21 signals in a flexible and efficient manner. The high efficiency
22 approach is applied to both combined-aperture, with C/A, P(Y), MA
23 and MT passed through the same upconverter amplifier chain and
24 antenna and separate-aperture, with MA and MT signals transmitted
25 out of a separate upconverter, amplifier chain and antenna from
26 C/A and P(Y)) communication path. The CASM waveform is a highly
27 efficient means of quadrature multiplexing new GPS military
28 acquisition and tracking signals with flexibility in adjusting

1 the relative power of combined signals without greatly altering
2 the power efficiency of the GPS waveform. The combined-aperture
3 CASM offer power-efficient constant-envelope GPS modernization
4 waveforms including the combination of C/A, P and M-code signals.
5 The generated M-code signals are mathematically equivalent to the
6 spatially combined signals.

7
8 CASM employs a subcarrier to phase modulated new M-code
9 signals on the same carrier as the current C/A and P(Y) ranging
10 codes. The constant-envelope CASM modulation is an evolution of
11 the constant-envelope subcarrier modulation used on SGLS and
12 other terrestrial and space systems to quadrature-multiplexed
13 systems. Unlike prior subcarrier modulation developments, the
14 CASM utilizes cross-product intermodulation terms as new ranging
15 communication signals. In the present invention, using
16 subcarrier modulation, cross product terms are interpreted as
17 signals and not as losses. Preferably, the M-Code signals are
18 generated by employing square-wave subcarriers to modulate the
19 new military ranging signals, due to ease of hardware
20 implementation, but CASM may be generated using both squarewave
21 and sinewave subcarriers for GPS modernization. These and other
22 advantages will become more apparent from the following detailed
23 description of the preferred embodiment.

Brief Description of the Drawings

Figure 1 is a block diagram of a digital baseband phase modulator subcarrier modulation implementation of quadrature product subcarrier transmitter that can be used for generating general quadrature product subcarrier modulation (QPSM) signals as well as particular coherent adaptive subcarrier modulation (CASM) signals.

Figure 2 is a QPSM spectral plot.

Figure 3 is a CASM spectral plot.

Detailed Description of the Preferred Embodiment

A detailed description of the preferred embodiment is described with reference to the figures using reference designations as shown in the figures. Referring to Figure 1, a quadrature product subcarrier modulation (QPSM) transmitter is used for transmitting in-phase code C_I , inphase data D_I , quadrature code C_Q , quadrature data D_Q , subcarrier code C_S and subcarrier data D_S signals originating from a code and data generator 10. The in-phase code and data signals are mixed by mixer 12, NRZ formatted by formatter 14, and power scaled by power factor 16 using mixer 18 providing the in-phase signal I_0 . The quadrature code and data signals are mixed by mixer 20, NRZ formatted by formatter 22, and power scaled by power factor 24

using mixer 26 providing a quadrature signal Q_0 . The mixers 12, 18, 20, 26, power factors 16 and 24, and formatters 14 and 22 may be integrated as a Q/I phase processor module.

In-phase and quadrature data signals D_I and D_Q are received by a data partition function generator 28 providing a data partition signal α . In-phase and quadrature code signals C_I and C_Q are received by a code partition function generator 30 providing a code partition signal β . The code partition signal β is mixed by modulo two mixer 32 with the subcarrier code signal C_s providing a mixed code signal communicated to another modulo two mixer 34 connected to another NRZ formatter 36. The data partition signal α is also mixed by modulo two mixer 38 with the subcarrier data signal D_s providing a mixed data signal communicated to the modulo two mixer 34 providing a composite digital modulated signal to the formatter 36 that provides an encoded subcarrier modulation signal to the mixer 48.

A frequency synthesizer 40 is used to generate necessary clock signals for the generator 10. The synthesizer 40 also generates a subcarrier clock signal communicated to a digital subcarrier generator 42 for providing a digital subcarrier signal that is scaled by a scaler 46 using a modulation index m 44 to provide a scaled digital subcarrier signal communicated to a mixer 48 that mixes the scaled digital subcarrier signal with the encoded subcarrier modulation signal from the formatter 36 to providing a modulated subcarrier signal communicated to a sine cosine processor 50 providing a sine subcarrier signal SS and a

1 cosine subcarrier signal CS. The generators 28, 30, 42,
2 formatter 36, processor 50, scaler 46, and mixers 32, 34, 38, and
3 48 may be integrated as a subcarrier digital baseband processor
4 module.

5
6 The cosine subcarrier signal CS is communicated to mixers 52
7 and 54 respectively receiving the I_0 signal and Q_0 signal and
8 respectively providing a cosine I_0 product signal and a cosine Q_0
9 product signal. The sine subcarrier signal SS is communicated to
10 mixer 56 and 58 respectively receiving the I_0 signal and Q_0 signal
11 and respectively providing a sine I_0 product signal and a sine Q_0
12 product signal. The cosine I_0 product signal from mixer 52 and
13 the sine Q_0 product signal from mixer 58 are subtracted by
14 subtractor 60 providing a digital rotated I signal. The cosine Q_0
15 product signal and the sine I_0 signal are added by an adder 62
16 providing a digital rotated Q signal. The digital rotated I
17 signal from the subtractor 60 is communicated to an in-phase
18 digital-to-analog converter (IDAC) 64 providing an analog rotated
19 I signal. The digital rotated Q signal from the adder 62 is
20 communicated to a quadrature digital-to-analog converter (QDAC)
21 66 providing an analog rotated Q signal.

22
23 The analog rotated I signal from IDAC 64 and the analog
24 rotated Q signal from the QDAC 66 are received by a conventional
25 quadrature modulator 68 providing an intermediate frequency (IF)
26 modulated signal by quadrature modulation using an IF local
27 oscillator signal IFLO from the frequency synthesizer 40. The IF
28 modulated signal from quadrature modulator 68 is modulated by a

1 radio frequency (RF) local oscillator signal (RFLO) using a mixer
 2 70 that provides an RF modulated signal communicated to a high
 3 power amplifier 72 that, in turn, provides an amplified RF
 4 modulated signal to a band pass filter 74 that, in turn, provides
 5 a filtered RF modulated signal to a transmitting antenna 76 for
 6 broadcast communication of a composite signal $S(t)$. The RF
 7 modulated signal from mixer 70 is expressed by the $S_0(t)Q/I$
 8 equation, $S_0(t)$ and P_{TF} equations when the modulation index m is
 9 equal to zero. The gain of the high power amplifier is expressed
 10 as the square root of the total power P_T . The communication
 11 channel including the modulator 68, mixer 70, high power
 12 amplifier 72, channel band pass filter 74 and antenna 76 are
 13 conventional means well known in the art.

14
 15 The new QPSM transmitter includes the addition of the
 16 subcarrier frequency generator 42 in the subcarrier digital
 17 baseband processor module and the scaler 44 to phase modulate the
 18 digital subcarrier. The mathematical equivalent of a phase
 19 rotation is constructed in the baseband by operating on the
 20 unrotated I_0 and Q_0 signals. This is achieved by transforming I_0
 21 and Q_0 to $I(t)$ and $Q(t)$. This complex-baseband approach can be
 22 implemented digitally in an application specific integrated
 23 circuit as a flexible means of subcarrier generation using the
 24 subcarrier processor 50.

25
 26 The preferred transmitter shown in Figure 1 can be
 27 implemented in whole or in part using equivalent analog means.
 28 For example, the phase modulation may also be performed on the

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1 subcarrier directly subject to the frequency limitations of an
2 equivalent analog phase modulator. In this equivalent analog
3 implementation, the encoded subcarrier modulation signal is
4 directly phase modulated onto the RF carrier RFLO by applying the
5 encoded subcarrier modulation signal to the input of an analog
6 phase modulator, not shown, having a modulation index n that
7 phase modulates the RFLO signal. In this equivalent analog phase
8 modulator case, the sine and cosine subcarrier processor module
9 and mixers 52, 54, 56, 58, as well as subtractor 60 and adder 62
10 are bypassed and the I_0 signal and Q_0 signals are inputs directly
11 into the quadrature modulator 68 connected directly to the high
12 power amplifier 72.

13
14 In a single subcarrier QPSM, the phase $\phi(t)$, is augmented by
15 the subcarrier signal, $\theta(t)$ modulated subcarrier signal from the
16 mixer 48.

17
18
$$\theta(t) = m\phi_s(t) = mD_s(t)\alpha_D(t)C_s(t)\beta_C(t)\phi_s(t)$$

19
20

21 The term $\phi_s(t)$ is a periodic subcarrier signal that may be
22 an arbitrary periodic waveform such as sine-wave, square-wave,
23 triangle-wave, having an angular frequency $\omega_s = 2\pi f_s$, in rad/sec.
24 The term m is the modulation index in radians. $D_s(t)$ is the
25 subcarrier data. $C_s(t)$ is a non-return-to-zero (NRZ)-encoded PRN
26 code with chipping rate R_{cs} . The term $\phi_s(t)$ represents the
27 composite subcarrier signal for the case of a single subcarrier.
28 $\alpha_D(t)$ and $\beta_C(t)$ are a function of the type of QPSM modulation

employed. These functions are typically binary functions and are used to preform the appropriate binary signals so that the desired resultant subcarrier code and data modulations will result in the composite spectrum. Using these definitions for forming $I_0(t)$ and $Q_0(t)$, the QPSM waveform is defined as $S(t)$. $S(t)$ represents the phase modulation of $S_0(t)$ by $\theta(t)$, and can be expressed by $I(t)$ in-phase and $Q(t)$ quadrature components out of the RF mixer 70.

$$s(t) = I_0(t) \cos(\omega_c t + m\phi_s(t)) - Q_0(t) \sin(\omega_c t + m\phi_s(t))$$

The $S(t)$ subcarrier modulated $S(t)$ RF signal from mixer 70 can be written in an expanded form.

$$\begin{aligned} s(t) &= \left\{ \begin{array}{l} I_0(t) \cos(m\phi_s(t)) \\ -Q_0(t) \sin(m\phi_s(t)) \end{array} \right\} \cos(\omega_c t) \\ &\quad - \left\{ \begin{array}{l} Q_0(t) \cos(m\phi_s(t)) \\ +I_0(t) \sin(m\phi_s(t)) \end{array} \right\} \sin(\omega_c t) \\ &= I(t) \cos(\omega_c t) - Q(t) \sin(\omega_c t). \end{aligned}$$

The analog $I(t)$ and $Q(t)$ signal, respectively from IDAC 64 and QDAC 66 are subcarrier rotated signals expressed as subcarrier rotated carrier modulations, $I_0(t)$ and $Q_0(t)$, as a function of subcarrier phase $\phi(t)$.

$$I(t) = I_0(t) \cos(m\phi_s(t)) - Q_0(t) \sin(m\phi_s(t))$$

$$Q(t) = Q_0(t) \cos(m\phi_s(t)) + I_0(t) \sin(m\phi_s(t))$$

The envelope $A(t)$ of the subcarrier augmented waveform is constant as the square root of the two times the sum of in-phase power P_I plus the quadrature power P_Q . Because the subcarrier modulation $\theta(t) = m\phi_s(t)$ is a phase modulation equivalent to rotation of both I_0 and Q_0 components of $S_0(t)$, the constant envelope is maintained independent of the choice of $\theta(t)$. Hence, the total average power envelope of the subcarrier augmented waveform remains constant and is the sum of the in-phase power P_I and the quadrature power, P_Q .

A single sinewave subcarrier can be an efficient means to modulate three signals in a constant-envelope waveform. For this case, $\phi_s = \sin(\omega_s t)$, so that $m\phi_s$ is a function of $\sin(\omega_s t)$.

$$m\phi_s = mD_s(t) \alpha_D(t) C_s(t) \beta_C(t) \sin(\omega_s t).$$

The $\cos(m\phi_s(t))$ and $\sin(m\phi_s(t))$ terms may be expanded in terms of Bessel functions. The sinewave $S(t)$ equation is a good approximate to terms of order J_2 .

$$\begin{aligned}
s(t) \cong & \left\{ \begin{aligned} & \sqrt{2P_I} \cdot J_0(m) D_I(t) C_I - \\ & \sqrt{2P_Q} \cdot 2J_1(m) D_Q(t) C_Q(t) \alpha_d(t) d_s(t) \beta_c(t) C_s(t) \sin(\omega_s t) \end{aligned} \right\} \cos(\omega_c t) \\
& - \left\{ \begin{aligned} & \sqrt{2P_Q} \cdot J_0(m) D_Q(t) C_Q(t) + \\ & \sqrt{2P_I} \cdot 2J_1(m) D_I(t) C_I(t) \alpha_d(t) d_s(t) \beta_c(t) C_s(t) \sin(\omega_s t) \end{aligned} \right\} \sin(\omega_c t) \\
& + \sqrt{2P_I} \cdot 2J_2(m) D_I(t) C_I(t) \cos(2\omega_s t) \cos(\omega_c t) \\
& - \sqrt{2P_Q} \cdot 2J_2(m) D_Q(t) C_Q(t) \cos(2\omega_s t) \sin(\omega_c t).
\end{aligned}$$

In the sinewave subcarrier QPSM approach, it is convenient to assume that the modulation index of the subcarrier waveform, m is less than $\pi/2$ rads. For this case, the power contained in the J_2 terms is much less than the power contained in J_1 and J_0 terms, for example, for $M = 1$ rad the power of J_2 term is less than 10db below the J_1 and J_0 terms. Furthermore, because these terms are spectrally separated from the carrier and subcarrier terms, the J_1 and J_0 terms will have a negligible contribution to the baseband correlations with $C_I(t)$ and $C_Q(t)$. Thus, neglecting the J_2 terms and applying the identities equations yields good approximation of the power content of $S(t)$ for a sinewave subcarrier QPSM.

The term $\alpha_D(t)$ is defined as the data partition function with $[\alpha_D(t)]^2=1$. The term $\beta_C(t)$ is defined as the code partition function with $[\beta_C(t)]^2=1$. Both α_D and β_C are defined in terms of QPSM variations. Both data and code QPSM messages have several variations. For $\alpha_D(t) = D(t) = D_I(t) = D_Q(t)$, the modulation is defined as doubly balanced QPSM because $D(t)$ is balanced on both carrier channels and $D_I(t) = D_Q(t) = D_s(t)$ is balanced on both subcarrier channels. In this case, the data modulation may be obtained by first tracking the carrier or the subcarrier signals. For $\alpha_D(t) = D_s(t)$, and $D_I(t) = D_Q(t) = D(t)$, the modulation is

1 defined as singly balanced QPSM because $D(t)$ is balanced on both
 2 carrier channels and the data modulation on the subcarrier
 3 signals is identical to that on the carrier signals. In this
 4 case, the data modulation may be obtained by first tracking the
 5 carrier or the subcarrier signals. For $\alpha_D(t) = D_s(t)$ and $D_I(t) \neq$
 6 $D_Q(t)$, the modulation is defined as, singly unbalanced QPSM
 7 because $D(t)$ is balanced on both carrier channels and the data
 8 modulation on the subcarrier signals is identical to that on the
 9 carrier signals. In this case, the data modulation may also be
 10 obtained by first tracking the carrier or the subcarrier signals.
 11 For $\alpha_D(t) = 1$, with $D_I(t)$, $D_Q(t)$, specified as different data
 12 messages, the modulation is defined as unbalanced QPSM because
 13 the data modulations on the in-phase and quadrature channels are
 14 different. For this case, the $D_I(t)$, $D_Q(t)$ messages must be
 15 demodulated prior to the demodulation of the $D_s(t)$ subcarrier
 16 message. For $\beta_C(t) = 1$, the resulting code is defined as normal
 17 QPSM. In this case, the carrier and subcarrier codes are
 18 selected so that the product is a higher order code. For
 19 example, the subcarrier code may be selected as a member of a
 20 preferred pair of M-sequences so that the resultant
 21 carrier-subcarrier product code is a gold code. For $\beta_C(t) =$
 22 $C_I(t)C_Q(t)$, the resulting code is defined as quadrature reversed
 23 QPSM. For $\beta_C(t) = C_s(t)$, the resulting code is defined as dual code
 24 QPSM. In general $\beta_C(t)$ may take on any combination of codes or
 25 functional forms to tailor the code shape and energy distribution
 26 of the QPSM waveform. Combinations of code and data variations
 27 are designated by first data designation followed by code
 28 designation followed by QPSM. For example, a combination of

1 $\alpha_D(t) = D_s(t)$ and $D_I(t) \neq D_Q(t)$ and $\beta_C(t) = C_I(t)C_Q(t)$ would be known as
 2 singly unbalanced, quadrature reversed QPSM.

3
 4 The QPSM waveform may also be developed for the squarewave
 5 subcarrier by setting $\phi_s(t) = \text{Sqr}(\omega_s t)$ For this case, $\text{Cos}(m\phi_s t) =$
 6 $\text{Cos}(m)$ and $\text{Sin}(m\phi_s t) = \text{Sin}(m)\phi_s(t)$. After applying these
 7 definitions, the QPSM $S(t)$ waveform for the square-wave
 8 subcarrier may then be expressed.

$$\begin{aligned}
 s(t) = & \sqrt{2P_I} \cdot \text{Cos}(m) D_I(t) C_I(t) \text{Cos}(\omega_c t) \\
 & - \sqrt{2P_Q} \cdot \text{Cos}(m) D_Q(t) C_Q(t) \text{Sin}(\omega_c t) \\
 & - \sqrt{2P_Q} \cdot \text{Sin}(m) D_{QS}(t) C_{QS}(t) \text{Sqr}(\omega_s t) \text{Cos}(\omega_c t) \\
 & - \sqrt{2P_I} \cdot \text{Sin}(m) D_{IS}(t) C_{IS}(t) \text{Sqr}(\omega_s t) \text{Sin}(\omega_c t)
 \end{aligned}$$

19 By appropriate choice of the modulation index m for sinewave
 20 QPSM, most of the usable power in the subcarrier waveform is
 21 transmitted to the signals, C_I , C_Q , C_{IS} and C_{QS} . The total power of
 22 $S(t)$ is expressed in terms of the total power of each component P_I
 23 and P_Q for each J_0 , J_1 , and J_2 harmonics. The wasted power in the
 24 higher order harmonics of $S(t)$ is unused and constitutes a signal
 25 loss. The power efficiency of the subcarrier waveform is then
 26 given by dividing the received power of the J_0 carrier and J_1
 27 subcarrier harmonics by the total power of $P_T = P_I + P_Q$, and
 28 expressed by a sinewave power efficiency equation.

$$\eta_p(m) = \frac{P_I * J_0^2(m) + P_Q * J_0^2(m) + P_I * 2J_1^2(m) + P_Q * 2J_1^2(m)}{P_I + P_Q} = J_0^2(m) + 2J_1^2(m)$$

The powers delivered to the carrier $P_c = P_I J_0^2(m)$ component and subcarrier first harmonic $P_s = P_I 2J_1^2(m)$ component can be compared. To select roughly equal power distribution between all four I and Q carriers, and I and Q subcarrier components, a modulation index of $m=1.16$ with $P_I = P_Q$ is selected. In this case, the efficiency of the sinewave subcarrier modulation approaches 95%. The power efficiency of the QPSM waveform treats the second and higher order subcarrier components as harmonic losses. For most applications, these components are outside the band of interest and/or carry too little energy to be detected. The QPSM waveform is over 95% power efficient, that is, only a 0.2 dB power loss, for modulation indices less than 1.16 for P_c less than or equal to P_{sc} . For values of m less than 1.2, the efficiency is over 95%. This efficiency, combined with QPSM constant-envelope makes the QPSM an effective modulation approach to augment existing quadrature-multiplexed communication systems or multiplex, spectrally separated communication channels about the same carrier.

For the case of a squarewave subcarrier, the total power of $S(t)$ may be expressed as a function of the P_I and P_Q components. When both I_s and Q_s subcarrier signals are tracked in the QPSM, the efficiency of the QPSM waveform is 100% for all m , assuming

negligible band-limiting losses. This is illustrated by the following squarewave power efficiency equation.

$$\eta_p(m) = \frac{P_I \cdot \cos^2(m) + P_Q \cdot \cos^2(m) + P_I \cdot \sin^2(m) + P_Q \cdot \sin^2(m)}{P_I + P_Q} = 1$$

In reality, however, only signals modulated to the first harmonic of the squarewave subcarrier are contained in the band of interest, with the bandwidth of the subcarrier signals being several times less than the subcarrier frequency. Band limiting effects can be approximated by employing the Fourier expansion of the squarewave signal to order $n=2k-1$, to obtain the approximate result for a band-limited channel. The power in the subcarrier is proportional to the RMS value of $\text{Sqr}(\omega_s t)$. A typical approximate made be realized for squarewave subcarriers with neglected overlap between respective code and data signals at squarewave harmonics in band-limited channels for a squarewave subcarrier modulated QPSM. The power efficiency of the square-wave subcarrier QPSM with the modulation index corresponding to equal carrier and subcarrier power is $m = 0.84$ and yields a power efficiency of 89% expressed by an approximate squarewave power efficiency equation.

$$\eta_p(m) = \cos^2(m) + \left(\frac{8}{\pi^2}\right) \sin^2(m)$$

1 Under realistic constraints of a band-limited channel, the
2 squarewave subcarrier is generally less power efficient than the
3 sinewave subcarrier.

4
5 Referring to Figure 2, a representative example of the QPSM
6 modulation is shown by the power spectral density (PSD) of a
7 singly-balanced QPSM presented for the case of a
8 quadrature-multiplexed direct sequence spread-spectrum system
9 with C_1 selected as a 2.0 MHz PRN code, C_Q selected as a 1.0 MHz
10 PRN code and C_s selected as a 4.0 MHz PRN code. In this example,
11 the data modulation rates are selected at 10.0 KBPS so that the
12 dominant spectral features would be the subcarrier modulated I_0
13 and Q_0 PRN codes. The subcarrier was selected as a 12.0 MHz
14 sinewave.

15
16 The generalized quadrature product subcarrier modulation
17 applies subcarrier modulation to quadrature-multiplexed
18 communication systems without the need to employ non-constant
19 envelope modulation subject to amplitude modulation to amplitude
20 modulation, and amplitude modulation to phase modulation
21 distortions through a nonlinear high power amplifier. The
22 spectral components reveal improved power distribution and power
23 efficiency for both the square-wave and sinewave subcarriers.
24 The power efficiency is approximately over 95% efficient for
25 sinewave subcarriers and over 90% efficient for square-wave
26 subcarriers for the case of equal carrier and subcarrier powers
27 in band-limited channels.

1 The resultant waveforms are for a single subcarrier,
2 however, QPSM can be applied to multiple subcarrier signals.
3 Employing multiple rates in the general case, product codes can
4 produce the desired spectral components in the resultant waveform
5 and these signals are generalized by defining the multiple
6 subcarrier phase modulation. In a generalized QPSM, a composite
7 baseband signal of n-subcarriers is formed by replacing $m\phi_s$ in
8 the original approach with the composite signal. The subcarrier
9 can be a single sinewave with $\theta(t)=\sin(\omega_s t)$ or a squarewave with
10 $\theta(t)=\text{Sqr}(\omega t)$, or other periodic functions. $\theta(t)$ is substituted
11 into the subcarrier modulation equation $m\phi_s$, to generate a
12 generalized subcarrier equation.

$$\theta(t) = \sum m^K D_s^K(t) \alpha_s^K(t) C_s^K(t) \beta_s^K(t) \Phi^K(t)$$

15 The term m^K is the modulation index of the K-th subcarrier.
16 In the generalized form of the invention, $\theta_K(t)$ is the K-th
17 subcarrier and may be any periodic function having radian
18 frequency, $\omega_s^K = 2\pi f_s^K$. D_s^K is K-th data sequence of the K-th
19 subcarrier. $\alpha_p^K(t)$ is the data partition function of the K-th
20 subcarrier and $\beta_c^K(t)$ is the code partition function of the K-th
21 subcarrier. These partition functions, generally need not be
22 binary.

24 QPSM employs the code and data partition functions to form
25 desired subcarrier harmonics that are used to transmit
26 information. Hence, cross-product terms are used as new
27 communication signals. That is, the QPSM pre-multiplies the
28

1 spread-data waveform by another spread data waveform so that the
 2 resultant product, naturally arising from the subcarrier
 3 modulation, is usable as a new communication signal. By doing
 4 so, the majority of the signals in the waveforms are designed for
 5 use as usable signals rather than allocated to wasted energy of
 6 intermodulation products. By pre-multiplying by code and data
 7 partition functions, QPSM allows for the implementation of
 8 subcarrier modulation on quadrature multiplexed communication
 9 signals. QPSM applies generally to any quadrature waveform
 10 including spread-spectrum waveforms. For the general case,
 11 neither the data nor the code partition functions must be binary.
 12 The functions need only to be selected such that the
 13 cross-product terms in the resultant subcarrier modulation are
 14 designed to be the signal of interest.

15
 16 A specific example of the general QPSM to quadrature spread
 17 spectrum communication waveforms is coherent adaptive subcarrier
 18 modulation (CASM). CASM allows for the presence of two
 19 time-varying quadrature multiplexed carrier signals in addition
 20 to the subcarrier signals that modulate the phase of the carrier,
 21 that are modulated producing a constant envelope composite signal
 22 using a single modulation index. The ability to employ
 23 subcarrier phase modulation to a quadrature-modulated signal is
 24 facilitated by the unique formulation of the CASM. CASM can be
 25 applied to GPS waveforms utilizing quadrature multiplexing and
 26 subcarrier modulation to transmit at least three GPS ranging
 27 signals in the same constant-envelope waveform. The specific
 28 formulation of the CASM is generalized to multiple civilian and

1 military GPS signals that may be on any combination of the $I_0(t)$
2 carrier, $Q_0(t)$ carrier and subcarrier signals. The formatting of
3 these signals is not a limitation of CASM and the partition
4 functions $\alpha_D^K(t)$ and $\beta_C^K(t)$ defined for the i -th subcarrier, and
5 may be composed from any combination of NRZ spreading codes,
6 other periodic function at any frequency, and the current GPS
7 data modulation commonly known as the GPS navigation message,
8 and/or spreading codes including the GPS C/A code, and P/Y code.
9 The specific CASM is selected to optimize the signals of interest
10 and to suppress interference between signals in the composite
11 waveform. Hence, the QPSM is not limited to sinewave subcarriers
12 but may be generalized to an arbitrary number of square-wave, or
13 other periodic, subcarrier signals. The CASM describes a
14 specific methodology for use of multiple code and data partition
15 functions to form an inherently constant-envelope waveform from
16 products of carrier signals, code and data partition functions,
17 and one or more subcarrier signals selected to optimize the
18 signals of interest and suppress interference between signals.

19
20 CASM is applied to the modulation of coherently related
21 quadrature duplexed military acquisition (MA) and military
22 tracking (MT) signals onto quadrature phases of a subcarrier in
23 constant envelope. CASM enables the modulation of new signals
24 onto the quadrature multiplexed GPS C/A and P(Y) code signals.
25 CASM pre-multiplies the subcarrier data and code modulation by
26 the binary data and code partition functions, α_D and β_C to enable
27 the utilization of subcarrier intermodulation products that would
28 otherwise be unusable signals. By so doing, the subcarrier

1 modulation results in a highly efficient waveform, with two
2 aspects of the same signal that may be optimized for military
3 acquisition MA and military tracking MT signals. The CASM
4 approach is based on the more general quadrature product
5 subcarrier modulation (QPSM) applicable to the augmentation of an
6 existing quadrature multiplexed communication system with new
7 communication signals.

8
9 In Figure 3, the power spectral density of the a
10 quadrature-multiplexed spread-spectrum transmitter is shown. The
11 transmitter uses a modulation index $m=1.16$ rad for a QPSM
12 waveform obtained for a complex-baseband implementation of the
13 output of a QPSM transmitter. The carrier modulation curve can
14 be thought of as a subset of the QPSM signals with $m=0$. The
15 spectrum illustrates the spectral components of the QPSM waveform
16 at \pm multiples of the 12.0 MHz subcarrier frequency. The
17 position and magnitude of the QPSM harmonics reveal harmonic
18 distribution and power content. In particular, the magnitude of
19 the J_2 harmonic of the carrier for $m=1.16$ is less than 10.0 dB
20 below that of the J_0 carrier components.

21 For the GPS CASM applications, α_p is equal to the GPS
22 navigation data message and $\beta_c(t)$ is equal to C/A or P/Y. In the
23 exemplar form, sinewave or square-wave subcarriers may be used.
24 For a sinewave subcarrier, $\theta_s = \sin(\omega_s t)$. For square-wave
25 subcarriers, one can generate Manchester encoded PRN code signals
26 by setting $\theta_s(t) = \text{Sqr}(\omega_s t)$, with the frequency of the squarewave
27 subcarrier signal set equal to the frequency of the PRN code
28 signal.

1 Sin(m) $\phi_s(t)$ identities for a binary subcarrier modulation phase
2 ϕ_s . Only the constant values Cos(m) and Sin(m) need to be
3 evaluated for the squarewave CASM.

4
5 CASM can be used with a single sinewave subcarrier as a
6 constant envelope multiple mode spread spectrum subcarrier
7 modulation for GPS. Considering only terms to order J2. In
8 general, there are an infinite number of carrier harmonics (J0,
9 J2, J4, ...) and subcarrier harmonics (J1, J3, J5...) but only
10 the first carrier and subcarrier harmonics need be considered for
11 $m \leq \pi/2$. For this range of the modulation index, the sinewave
12 CASM approach is over 95% power efficient. For values of m
13 greater than π/w , a much greater fraction of the transmitter
14 power is lost to the n=2, 3, 4, 5 harmonics outside of the GPS
15 band depending upon subcarrier selection. The current GPS
16 waveform is recovered by setting m=0. The composite subcarrier
17 PRN product codes and data messages can be expressed to a good
18 approximation by only the dominant terms in the harmonic
19 expansion by applying the sinewave subcarrier expression with the
20 use of some well-known Bessel function identities. By the
21 selection of $\alpha_D(t)$ and $\beta_D(t)$, C_{IS} and C_{QS} are coherently related to
22 new military acquisition and tracking signals with the same
23 M-code data modulation on both of these subcarrier signals. With
24 the current partitioning of subcarrier code and data, the new MA
25 acquisition and MT tracking codes are derived from the single new
26 military PRN code $C_s(t)$. The P(Y)-code is a very long PRN code
27 repeating each week. The randomness properties of the P(Y) code
28 are used by military GPS receivers. By selecting the C_s code to

1 be a non-periodic tracking code, the correlation randomness
 2 properties of the MT code will be determined primarily by the
 3 properties of the C_s code. The appropriate selection of the C_s
 4 code relatively prime to the product of (C/A-code)*(P(Y)-Code)
 5 will result in a code with a length that is equal to the product
 6 of the lengths of the C_s and the P(Y)-Codes, much longer than the
 7 current P(Y)-code. The generation of a long code from two
 8 relatively prime sequences is well known and is the primary basis
 9 for the current GPS P(Y)-code length. The M-code data modulation
 10 is determined by the current selection of data partition
 11 function, similar to the case of the PRN codes, by $D_M(t) = D_{IS}(t)$
 12 $= D_{QS}(t) = D_s(t)$ where $D_M(t)$ is defined as the new military
 13 navigation data message. The current selection of the data
 14 partition function leading to a single M-code data message, is
 15 convenient to adapt the CASM to the currently favored option of a
 16 single new M-code data modulation for the new military M-code
 17 signals but is not a limitation of the general approach. By
 18 setting $\beta_C(t) = C_P(t)$ instead of the C/A code, the I/Q phasing of
 19 the MA and MT codes are reversed. For this case, $C_{MA} = C_{IS}(t) =$
 20 $C_s(t)$ and $C_{MT}(t) = C_{QS}(t) = C_s(t)C_{C/A}(t)C_P(t)$. For this case, the
 21 partition of the data modulation remains unchanged and the MA and
 22 MT signals obtain power from alternate phases of the carrier,
 23 that is, the MA signal on the Q channel and MT signal on the I
 24 channel rather than the MA signal on I channel and the MT signal
 25 on the Q channel. The $C_s(t)$ is defined as $C_{MA}(t)$. A single new
 26 subcarrier acquisition code $C_s(t)$ is defined to generate both an
 27 acquisition and a tracking code signal. The designation
 28 acquisition and tracking does not mean that these signals cannot

1 both be acquired and tracked, analogous to the case of C/A and
2 P(Y) codes, but only that these signals are optimized for the
3 these GPS operating states. The efficient use of this code with
4 CASM utilizes this code as a useful ranging signal in a well
5 controlled manner. The power contained in the MA and MT signals
6 can be combined using a dual-code correlator to utilize the total
7 power in the MA and MT signals.

8
9 Due to the ease of digital implementation, square-wave
10 modulated ranging codes have become popular for use as a new
11 M-code signal. The resulting higher harmonics of the carrier are
12 not present in the square-wave CASM approach. Thus, to raise the
13 subcarrier power well above the carrier power, M is increased to
14 a value near $\pi/2$. For this case, the only losses are due to the
15 band-limiting effects inherent in any square-wave modulated
16 ranging signal. The current GPS waveform can be recovered for the
17 case of $m=0$.

18
19 The sinewave and square-wave CASM approaches may be
20 generalized to an arbitrary number of subcarriers into the
21 generalized modulation subcarrier equation. This method may be
22 useful to enable multiple military and, perhaps, additional
23 civilian ranging signals to be transmitted in a single constant
24 envelope waveform. The notation CASM (r,n) denotes the military
25 acquisition (MA) and military tracking (MT) pair of signals
26 generated by the CASM modulation method. The MA signal, MA
27 (r,n), is formed from the product of a $n \times 1.023$ MCPS spreading
28 code modulated by a $r \times 1.023$ MHz subcarrier and the MT signal,

1 MT(r,n), is given by the MA signal modulated by C/A code and P(Y)
 2 code. In general, the MT code is given by the product of the MA
 3 code times an unencoded 10.23 MHz NRZ code. The MT-code signal
 4 will have a length greater than or equal to the current P(Y)-code
 5 signal, depending upon the selection of the MA code.

6
 7 As an example of the CASM signals, a CASM(10,2) quadrature
 8 modulated pair of ranging signals consists of a MA(10,2)
 9 acquisition signal on the I phase of the carrier along with the
 10 P(Y) code and an MT(10,10) tracking signal on the Q phase along
 11 with the C/A Code that are modulated using the modulator depicted
 12 in Figure 1 with C_I equal to the P(Y) code, C_Q equal to the C/A
 13 code and C_s equal to a 2.046 MCPS NRZ formatted M-Code. In this
 14 example, the subcarrier signal is a 10.23 MHz squarewave, D_I is
 15 equal to D_Q that is also equal to the current GPS data navigation
 16 message and D_s is the new military GPS subcarrier data message.
 17 The data partition function, α is equal to the current GPS
 18 navigation message and β is equal to the C/A code.

19
 20
 21 The power spectral densities (PSD) of this example are
 22 compared to the composite P(Y) code and C/A code spectrum
 23 depicted in Figure 3 for the MA(10,2) and MT(10,10) signals
 24 respectively. With the exception of the P(Y)-code PSD, which was
 25 calculated analytically, the MA and MT code PSDs were determined
 26 from a simulation employing the exact C/A-codes and P(Y)-codes
 27 for PRN 1, and a 20'th order M-sequence for the MA Code. The
 28 PSD's were then obtained from a 1024 point Fast Fourier Transform

1 followed by a discrete time autocorrelation using 512 Lags and a
2 Hanning window function. The sampling rate of the simulation was
3 1.2276×10^8 samples per second. The MA signal for this option
4 consists of a 2.046 MHz PRN code ($C_s(t)$) modulated by a 10.23 MHz
5 squarewave subcarrier. The product of $C_s(t)C_p(t)C_{CIA}(t)$ results in
6 an unbalanced 10.23 MHz PRN Code. The product of this code with
7 a squarewave modulated ranging signal results in a 10.23 MHz
8 Manchester code.

9
10 In general, it can be shown that the integer relation
11 between the subcarrier frequency and the PRN code chip rate, as
12 in the CASM(10,2) example, is not necessary for new M-code
13 signals to have minimal interference with the C/A-code and P(Y)-
14 code signals from a given satellite. Rather, it is only
15 necessary that the frequency of the subcarrier, such as a
16 sinewave or squarewave, is much greater than the GPS data rate.
17 This is true for any subcarrier of frequency greater than 1.023
18 MHz.

19
20
21 The non-integer relationship has been simulated for the case
22 of a CASM using a 5.115 MHz sinewave subcarrier modulating a
23 4.092 MHz PRN code (C_s). For this example, the simulated variance
24 of the on-time C/A code correlated with the received Q-signal
25 ($C/A + MT$) and on-time P(Y)-code correlated with the received I-
26 signal ($P + MA$) was 4.14×10^{-5} (with unit power in the PRN-codes).
27 Relative to a signal level of -160 dBW, this value corresponds to
28 a multiplexing noise density of only -203 dBW as compared to

1 about -140 dB for thermal noise power out of a C/A-code
2 correlator. This level is similar to the cross-correlation
3 magnitude observed for the correlation of on-time C/A-code with a
4 filtered 5.115 MHz Manchester code. This magnitude has been
5 postulated to be identically zero if the bandwidth of the GPS
6 system were infinite. This performance, together with CASM's
7 inherent constant-envelope nature, make the approach a flexible,
8 efficient modulation technique, capable of accommodating a wide
9 range of signal modulation alternatives and providing the
10 capability to control spectral isolation between signals inside
11 and outside of the GPS band. Its flexibility minimizes the cost,
12 complexity and risk associated with implementation of the GPS
13 transmitter circuitry.

14
15 In the CASM waveform approach, both MA and MT code signals
16 are generated from a single new ranging code. Each of these
17 signals can be acquired and tracked independently of each other
18 and independently of the C/A and P(Y) codes. Because the MA and
19 MT ranging signals have low cross-correlation, as has been
20 verified through simulation, and these signals are coherently
21 related. The power of the MA and MT codes can be combined through
22 the implementation of a non-coherent dual code correlator, not
23 shown. The significant difference between the conventional
24 correlator and the dual code correlator is that the sum of the MT
25 and MA codes are used as local references rather than just the MA
26 and MT codes alone. This straightforward implementation will
27 result in the utilization of the total power contained in both
28 the MA and MT signals during acquisition and tracking.

1 The CASM approach enables a highly efficient means to
2 generate complementary military acquisition and tracking signals
3 that are adaptable to both separate and combined-aperture
4 implementations of the GPS modernization signals. In particular,
5 the QPSM-based CASM modulator permits the constant envelope
6 modulation of these new signals and existing GPS ranging signals
7 to enable efficient transmission through a single high-power
8 amplifier. Both sinewave and squarewave subcarrier
9 implementations of CASM are straightforward complex-baseband
10 implementation of a CASM modulator design. The general efficiency
11 of the approach for both squarewave and sinewave subcarriers was
12 demonstrated be in the range of 70% to 95% for practical
13 subcarrier power ratios. Those skilled in the art can make
14 enhancements, improvements and modifications to the invention,
15 and these enhancements, improvements and modifications may
16 nonetheless fall within the spirit and scope of the following
17 claims.